Recent Advances in Finline Circuits*

SLOAN D. ROBERTSON†

Summary—Detailed studies of transmission properties of the finline coupler have revealed existence of certain phenomena which adversely affect the transmission characteristics. A discussion of the origin and successful elimination of such effects is described.

New applications for the finline coupler have been found in the design of hybrid junctions, twists and bends in multimode waveguide, and other polarization selective devices.

The paper concludes with a presentation of theoretical and experimental results in the application of finline techniques to low-pass microwave filters.

With the demand for ever wider frequency bands in microwave communication systems, it is evident that means for broadbanding various circuit elements must be found. Finline techniques already offer the means of accomplishing this in several devices and may be expected to be of use in others.

Some finline circuits are capable of operating over at least a three-to-one frequency range. One of these, the basic finline coupler, announced last year,[1] is shown in Fig. 1. It consists of a length of circular waveguide

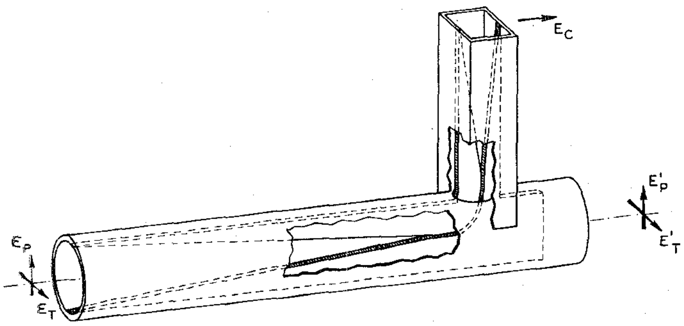

Fig. 1—Finline coupler.

fitted with a pair of diametrically opposite, thin metal fins which taper in from the outer wall of the guide until their opposing edges are separated by a narrow gap or slot at the center. Thus, substantially all of the energy associated with the electric field E_p (where the subscript p denotes that the vector is parallel to the plane of the fins) is transformed from the dominant mode of propagation in the circular waveguide to a finline mode in which the energy is largely confined to the gap and its immediate vicinity. This energy may then be removed from the circular guide by curving the finline and bringing it out through a small hole in the side wall. It may then be launched into another waveguide as shown.

On the other hand, a wave characterized by the transverse field E_T will pass on through the guide, relatively undisturbed by the presence of the fins, and will emerge as E_T'.

One can readily see that the finline coupler offers a means whereby it is possible to separate two waves perpendicularly polarized to one another. The fact that this is done with smooth tapers several wavelengths long suggests that the coupler ought to work over very wide bands. Such has been found to be the case.

If only one wave is present in the guide, it is found that one may abstract any desired proportion of its energy by simply rotating the coupler about its axis so that the plane of the fins makes an angle with respect to the plane of polarization of the wave. The abstracted field will then be proportional to the cosine of the angle.

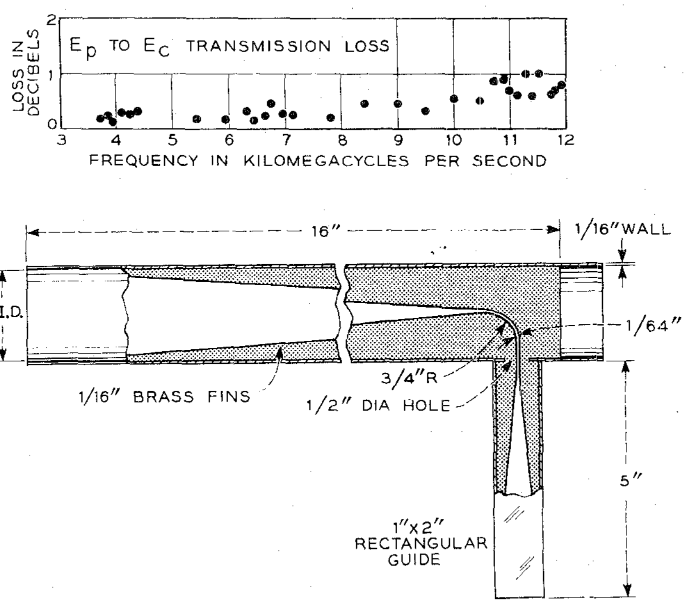

Fig. 2—Transmission characteristics of a coupler having the dimensions shown.

Fig. 2 shows a plot of the transmission losses from E_p to E_c for a coupler having the dimensions shown in the lower part of the figure. It will be observed that, for the most part, the losses are less than one db over the entire range of frequencies from 3.75 to 12.0 kmc. Over most of this range the losses are only a few tenths of a db. The losses encountered by the transverse mode in going from E_T to E_T' are less than 0.1 db, and are not plotted in the figure.

Mode Resonances in the Finline Coupler

Recent detailed transmission measurements on the finline coupler have uncovered several sharp peaks in

* Manuscript received by the PGMTT, July 16, 1956. Presented before the National Symposium on Microwave Techniques, Philadelphia, Pa., February 2–3, 1956.
† Goodyear Aircraft Corp., Litchfield Park, Ariz.
[1] S. D. Robertson, "The ultra-bandwidth finline coupler," Proc. IRE, vol. 43, pp. 739–741; June, 1955.

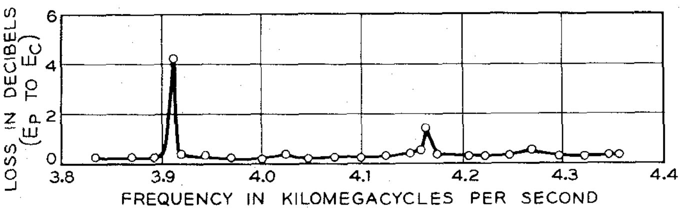

Fig. 3—Half-cylinder mode resonant loss peaks in the 4 kmc band.

the transmission loss curves which have not been observed in the earlier measurements of Fig. 1. A sample transmission loss curve showing two of these peaks in the 4 kmc band is shown in Fig. 3. It will be noted that the higher frequency peak is much smaller than the lower frequency one. In fact, careful measurements with swept frequency oscillators covering the bands between 5.8 and 6.6 kmc and also between 10.5 and 12.1 kmc have not shown any of these sharp peaks at all.

Studies have shown that these loss peaks are due to the excitation of a dominant mode in each of the half-cylinders separated by the fin. The field configurations of these dominant modes are sketched in the end view of the coupler in Fig. 4. The dominant modes are very

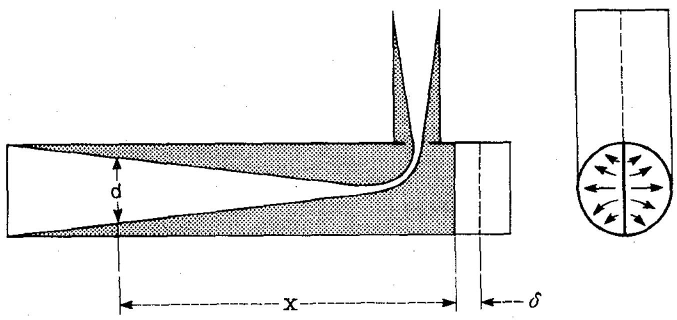

Fig. 4—Geometry of half-cylinder resonant modes.

likely generated in the vicinity of the curved portion of the finline. The geometry causes them to be 180° out of phase in the two half-cylinders. After being launched, the modes are propagated in both directions along the guide. Those propagated to the right soon encounter the edge of the fin and are reflected, since the vectors in each half-cylinder are out of phase and cancel beyond the fin edge. Likewise, the waves propagated to the left ultimately reach a point where the fin separation d is sufficient to permit the vectors to cancel and reflect the waves. If the spacing x between these two reflecting points is an integral number of half wavelengths, the sections will resonate and will absorb some of the useful power from the finline. This is believed to be the origin of the loss peaks in the curve.

These peaks will not all be equally objectionable. If the point where the finline launches the half-cylinder modes is at a position in the standing wave pattern of the resonant section where there is a favorable coupling between the finline and the resonant modes, the loss peaks may be rather high. At other frequencies the coupling may be unfavorable and the loss peaks may be low or even negligible. The absence of these peaks at the higher frequencies is possibly due to two factors. The increase of the electrical length of the tapers with frequency results in a better impedance match between the guide and the finline gap and lowers the coupling to the half-cylinder modes. On the other hand, increased ohmic losses at the higher frequencies would be expected to dampen the resonant peaks.

The identification of the loss peaks of Fig. 3 with half-cylinder mode resonances was easily confirmed. It was found that the locations of the peaks could be shifted by extending the back edge of the lower fin by a distance δ as shown in Fig. 4. With the fin extended, the length of the resonant chamber becomes $x+\delta$, and at resonance it must include an integral number of guide half-wavelengths.

$$x + \delta = n\lambda_g/2. \tag{1}$$

With the well-known relation between λ_g and the cut-off frequency f_c, (1) can be solved for the resonant frequency.

$$f = [n^2c^2 + 4f_c^2(x + \delta)^2]^{1/2}/2(x + \delta). \tag{2}$$

Observations of the shift in resonant frequency for three different, known values of δ made it possible to solve for the three unknowns x, n, and f_c. For the coupler of Fig. 2, x is found to be 27.6 cm, n is 3 for the 3.92 kmc peak, and f_c is 3.54 kmc. The latter is slightly higher than the theoretical cut-off frequency of 3.47 kmc for the half-cylinder mode in 2 inch diameter guide. This is due to the reduction in half-cylinder cross section caused by the thickness of fins. It is perhaps of interest that the fin separation d at the left-hand reflection point is about 62 per cent of the waveguide diameter.

Resonant peaks have been successfully eliminated by introducing loss at a point where it can absorb energy from the half-cylinder dominant modes. Fig. 5 shows the results of one experiment in which this was done. A 12.5 ohm per square resistive strip was placed in the plane of the fin adjacent to the right-hand edge. This strip absorbs energy from the half-cylinder modes which would be reflected at this point and kill the resonances. Since it lies in an equipotential plane of the transverse mode, the resistive strip will have little effect upon the transmission from E_T to E_T'.

Applications of Finline Couplers

Fig. 6 shows how two finline couplers may be arranged to form a hybrid junction. When the planes of the fins are inclined at an angle of 45 degrees, as shown in the figure, one obtains a 3 db hybrid. A wave entering at E_1 passes through the left-hand coupler without modification. Upon entering the second coupler, however, it is split into two equal components, one of which emerges at E_3, and the other emerges at E_4. Likewise, a wave entering E_2 travels through the first coupler and is split by the second coupler into two equal components. Degrees of coupling other than 3 db may be obtained by inclining the fins at angles other than 45 degrees.

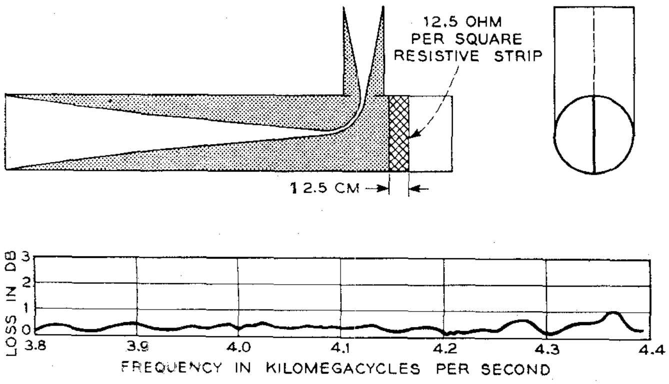

Fig. 5—Elimination of mode resonances.

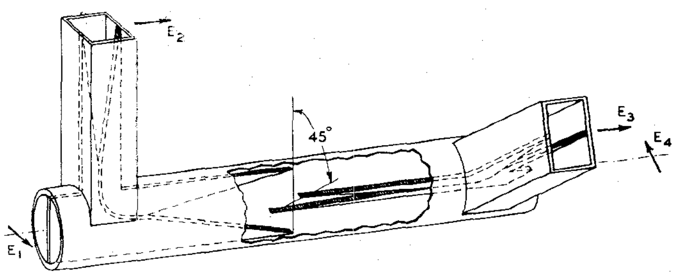

Fig. 6—Ultra-broadband hybrid junction using two finline couplers.

Fig. 7 shows another application of finline techniques to form a flexible waveguide joint or bend. As seen in the figure, the device comprises input and output waveguide sections containing 8-inch long tapers for matching in and out of the flexible finline section between. The fins in this case are exposed to free space in the region between the waveguide sections. It has been found that, for the device shown, the transmission losses are reasonably low in the 4 kmc band even when the flexible fins are bent or twisted. Lower losses are obtained by placing a sheet of dielectric between the fins in the exposed portion of the path. Apparently the action of the dielectric in confining the fields more closely to the gap, thereby reducing the radiation losses, exceeds the added dissipation due to the dielectric. The dielectric has been tapered at the ends for purposes of impedance matching. In the lower part of the figure a curve of the insertion loss over the 4 kmc frequency band is plotted. The joint can be bent or twisted through angles of as much as 45 degrees without substantial change in the transmission losses.

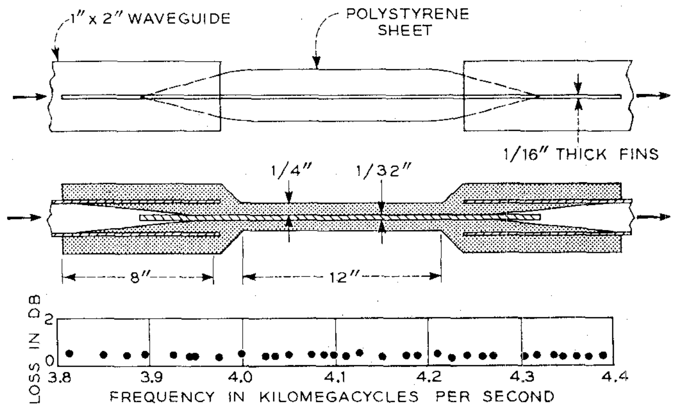

Fig. 7—Flexible finline bend or twist. Curve shows loss vs frequency.

It has been suggested that four finline couplers can be combined as shown in Fig. 8 to form a broad-band, right-angled, waveguide bend. The figure is of a model and is for illustrative purposes only. The large wooden cylinders represent hollow metal waveguides, and the brass strips represent finlines. It can be seen that the vertical component of a wave entering the lower waveguide will be collected by a finline coupler and will emerge through the side of the guide along the finline path shown and can then be launched into the upper waveguide by means of another coupler. Similarly, the horizontal component of the input wave will be collected by the horizontal coupler and will then be launched into the upper guide. It will be noticed that the arrangement possesses complete symmetry, and that the path lengths traveled by the two components are equal. Hence, there will be no phase difference between the two components as they emerge from the upper guide. Such a bend is expected to operate over the same broad frequency band over which the individual couplers have been found to operate.

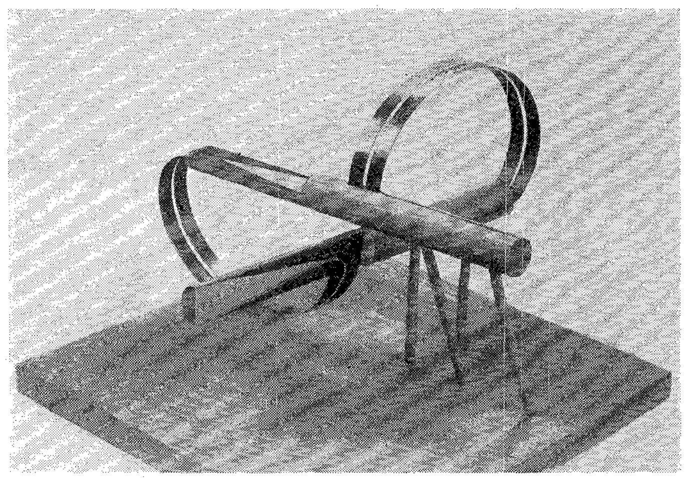

Fig. 8—Model showing the arrangement of four finline couplers in a broad-band, right angle bend in circular waveguide.

Finline Filters

With finline techniques it has been found possible to design frequency filters using the ordinary lumped-circuit theory that is used in low-frequency filter design. Analogs have been found which correspond to the series or shunt inductors and capacitors required for building up conventional π or T networks. Some of these analogs are shown in Fig. 9. It is seen that the finline equivalents for series inductance and shunt capacitance are quite simple, consisting only of a slot less than $\lambda/4$ in depth for the former, and a short tab protruding into the gap for the latter.

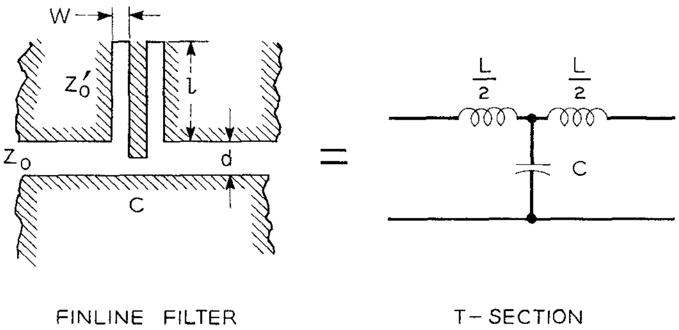

Fig. 9—The equivalence between a finline circuit and a low-pass T section.

Evidently, with a series inductor and a shunt capacitor, one may design a low-pass, T-section filter having a specified iterative impedance Z_0, a cut-off frequency f_c, and a frequency of infinite attenuation f_∞ corresponding to the frequency at which the series slots become antiresonant.

The standard low-pass T-section design equations are

$$f_c = 1/\pi(LC)^{1/2} \tag{3}$$

and

$$Z_0 = (L/C)^{1/2}[1 - (f/f_c)^2]^{1/2}. \tag{4}$$

In the usual filter, L and C are both constants. In the present case, however, C is a constant, but L varies with frequency according to the relation

$$L = (Z_0'/\pi f) \tan \beta l \tag{5}$$

where $\beta = 2\pi f/c$ and Z_0' is the characteristic impedance of the stub lines. Since L is a function of f, the value for f_c in (3) also varies with f. This means that, at any particular frequency below the true cut-off frequency, f_c in (3) gives the apparent cut-off frequency for the particular value of L then prevailing. The true cut-off f_c' is the particular value of f_c defining the point where Z_0 passes from a real to an imaginary value.

The variation of L with f leads to a similar difficulty in (4) in that Z_0 approaches zero as the frequency decreases. This can be evaded, however, in the microwave case since the frequencies of interest will not approach zero but will have a lower bound defined by the low-frequency cutoff of the finline waveguide itself. The problem may be resolved quite easily by defining the lowest operating frequency of interest f_1 and designing the filter so that it will exhibit the correct value of Z_0 at that frequency.

If f_c' is substituted for f in (3) and (5), one may combine these equations and solve for C. Likewise, if f_1 is substituted for f in (4) and (5), one may obtain a second expression for C. If the equations for C are then equated one obtains

$$Z_0'^2 = f_1 Z_0^2/[f_c' \tan \beta_c' l - f_1 \tan \beta_1 l] \tan \beta_1 l \tag{6}$$

where $\beta_c' = 2\pi f_c'/c$ and $\beta_1 = 2\pi f_1/c$.

It will be apparent that, having specified values for f_1, Z_0, f_c', and f_∞, one may immediately obtain $l = c/4f_\infty$, and then Z_0' the characteristic impedance of the stub line, from (6). The shunt capacitance C is given by

$$C = 1/\pi f_c' Z_0' \tan \beta_c' l. \tag{7}$$

Finally, it is necessary to determine the slot widths d and w to produce the required line impedances Z_0 and Z_0', respectively, and to compute the spacing between the stubs and the fin spacing in this region to establish the value of C.

S. P. Morgan[2] has dervied formulas for the characteristic impedance of a finline contained in a circular waveguide of a given radius. Numerical values have been obtained for the case of a circular guide of one inch radius with 1/16 inch thick fins separated by a gap spacing equal to d. For this particular fin thickness and for gaps ranging in value from 0.005 inch to 0.032 inch the characteristic impedance has been found to be given approximately by the formula,

$$Z_0 = 8d^{2/3} \qquad (d \text{ in mils}). \tag{8}$$

For finlines operating at frequencies far above cutoff, it is assumed that the characteristic impedance depends primarily upon the fin thickness and the gap spacing, and that it is relatively independent of the size or shape of the outer waveguide envelope. If this is true, one may use the above formula for calculating the required slot widths d and w in a finline filter irrespective of the form of the outer shell.

In order to test the general aspects of the theory, an experiment a filter has been designed and constructed. The specified parameters were $Z_0 = 45.5$ ohms, $f_1 = 3.0$ kmc, $f_c' = 8.5$ kmc, and $f_\infty = 11.05$ kmc. The computed quantities were $Z_0' = 25.4$ ohms, $C = 5.5 \times 10^{-13}$ farads, $d = 0.0135$ inch, $W = 0.0056$ inch, and $l = 0.268$ inch. The spacing between the stubs and the fin spacing in this

[2] Unpublished work.

region, required to establish C, were computed using the parallel plate expression for capacitance and making due allowance for fringing. These dimensions are shown in Fig. 11.

The variation of iterative impedance with frequency for this filter design has been computed and is shown in the curve of Fig. 10. It will be noted that the impedance is quite constant from 3 to 7 kmc. The impedance of a simple T-section filter, whose inductance L is constant with frequency, drops off much more rapidly with frequency than the curve given in the figure. The latter has more of the shape characteristic of the m-derived sections of filter theory.

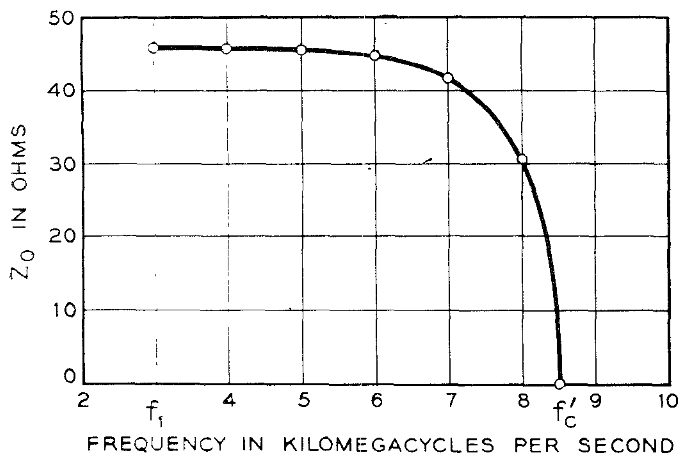

Fig. 10—Variation of the theoretical iterative impedance of a finline filter with frequency.

An experimental filter was built in conformity with the above design. Details of the filter are shown in Fig. 11. The fins were mounted in a section of 1 inch×2 inches rectangular waveguide. The ends of the fins, although not shown in the figure, were tapered in order to provide an impedance match to the waveguide.

The measured loss as a function of the frequency is shown in the lower part of the figure. The dotted portions of the curve represent frequencies which were not within the range of the measuring equipment used. Evidently the lumped-circuit design theory is capable of giving reasonably good results.

The results of these experiments show that it is possible to use finline techniques in building microwave filters. It is thought that a more detailed study of

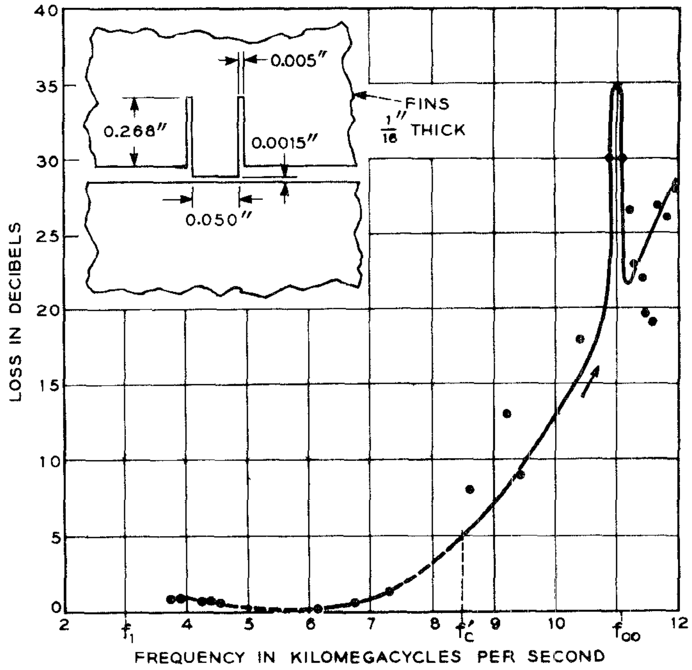

Fig. 11—Experimental transmission characteristics of a finline filter.

these techniques will lead to refinements resulting in filters having lower losses and probably sharper cutoffs than those measured with the exploratory model described here. Higher discriminations can probably be obtained by using several T sections in tandem.

The principles outlined here may possibly be applied to other wide-band microwave circuits such as equalizers, impedance transforming networks, etc.

Conclusion

Apparently finline techniques offer the means of designing certain microwave circuit components which will operate over much wider frequency bands than heretofore possible with standard components. It is thought likely that refinements and extensions of this work will lead to improved performance and unfold new applications of finline techniques.

Acknowledgment

The author wishes to express his appreciation to C. F. Chapman for his enthusiastic assistance in the experimental portions of this work.